A New Multipath and Noise Mitigation Technique Using Data / Data-less Navigation Signals

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BIOGRAPHIES

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ABSTRACT

Modernized GPS and Galileo will be available to civil users at the end of this decade. Their signal structure design will bring tremendous improvement for any user compared to current GPS. One important advance, which is the focus of this paper, is the availability of data and pilot (or data-less) channels wrapped in a Quadrature Phase Shift Keying (QPSK) modulation (or similar). The presence of a pilot channel will allow for a more effective tracking of the carrier, bringing lower loss of lock threshold, better resistance to dynamics, and better carrier aiding. A new method, using both the data and pilot channel, is developed herein that reduces the noise of the carrier-phase estimation while keeping the excellent performance of coherent carrier tracking. The use of a more reliable and accurate Phase Lock Loop (PLL) allows for longer integrations that can significantly aid the Delay Lock Loop (DLL) in mitigating thermal noise in case of a low signal-to-noise ratio. This possibility can also be used to integrate multipath mitigation techniques that usually have a lower resistance to ambient noise. A method using the High Resolution Correlator technique on the pilot channel and a Narrow Correlator™ technique on the data pilot is investigated. It shows excellent resistance to noise and multipath, while being very robust. Finally, the combination of the two carrier phase and code delay tracking using the data and pilot channels is tested in different environments including the presence of strong multipath. Static and kinematic users are simulated as well.

INTRODUCTION

With the forthcoming modernized GPS and Galileo, Global Navigation Satellite Systems (GNSSs) will try to meet the ever increasing demands of civil users. Many issues experienced by the current GPS have been considered in the design of these two new GNSSs. As an example, the availability of multiple civil frequencies will allow simple and efficient ionospheric modeling for most users. However, one source of error, because of its environment-dependent nature and its potentially high magnitude, remains a significant problem: multipath. An efficient way to reduce it has been used in the design of some of the new signals by increasing the spreading code rate (on GPS L5, and Galileo E5A and E5B). For lower spreading code rates such as for GPS L1 C/A, (but it could be extended to GPS L2c, Galileo E1), it can still be
a source of very high levels of errors using traditional tracking techniques. Many multipath mitigation solutions have been developed in the past 20 years at different stages of the receiver: (1) the antenna through an efficient design of the gain pattern, (2) the navigation module through the rejection of blunders using Receiver Autonomous Integrity Monitoring (RAIM) or other types of residual checking, and finally, (3) the tracking loops which will constitute the main core of the present research. Solutions at the tracking loops level have an edge on multipath mitigation techniques as, theoretically, multipath can come from anywhere and thus go through the antenna without being significantly affected, while residual checking is greatly dependent upon the measurements and redundancy obtained from the tracking loops themselves. As a result, tracking loops are a tremendously important stage to mitigate multipath. Several innovative ways of dealing with this problem have been developed. When considering pure multipath rejection without other sources of errors, and a slow code chipping rate, it appears that techniques such as the High Resolution Correlator (HRC) [McGraw and Braasch, 1999], or Strobe and Edge [Garin et al., 1996; Garin and Rousseau, 1997] have the best performance. However, they have an intrinsic robustness problem that makes them unsuitable for certain applications due to their discriminator shape and lower resistance to high noise levels [Van Dierendonck and Braasch, 1997].

One way to resolve this noise mitigation problem is to extend the pre-detection integration time in order to increase the post-correlation signal-to-noise ratio and have a better discrimination of the error. However, the existence of data on the current GPS C/A-code limits the coherent integration time to 20 ms. If a longer integration is needed, non-coherent integration, affected by potential squaring losses can be done. However, two main constraints intervene when long integrations are considered: first, the estimation of the Doppler has to be accurate in order to keep constructive correlation. Second, an integration time that is too long will have an impact on the tracking loops as the linear model for the Phase Lock Loop (PLL) and the Delay Lock Loop (DLL) assumes that the product of the filter loop bandwidth and the integration time is close to zero.

These problems can be partially contained when considering implementation on most of the future signals of a (Quadrature Phase Shift Keying) QPSK modulation with data-less (or pilot) and data channels. The existence of a pilot channel will allow the use of more efficient carrier tracking techniques that will strongly outperform the commonly used Costas loop. As a result, it will allow for better Doppler removal and in more constructive long correlations. Moreover, the use of two channels can lead to different combinations of discriminator outputs in order to try to reach an optimal noise mitigation. This study will constitute the first part of this research.

Once several carrier tracking techniques are investigated, code tracking will be studied. The Narrow Correlator™ and HRC method will be introduced for BPSK(1) and BOC(1,1) modulations before showing possible implementations of data/pilot DLL using these techniques. Finally, results from simulated data will be shown implementing the methods previously introduced. These simulations include hard multipath environment in static and dynamic conditions.

CARRIER TRACKING

Carrier tracking is very important in a GNSS receiver, as it allows for the estimation of the input signal phase which is essential to demodulate the navigation data. It also allows an estimation of the Doppler, thus giving an indication of the user’s dynamics. However, in current GPS, it is very often the first cause of a loss of lock due to its reduced resistance to external sources of errors such as thermal noise, oscillator phase error, oscillator vibration or dynamics. The availability of a pilot channel on most of the future GNSS signals will make it more robust by offering the possibility to use coherent carrier tracking. In this section, different discriminators that could be used for data and pilot channels are first presented. A detailed overview of the main sources of errors impairing carrier tracking is then presented, before introducing a new method to better track the carrier phase using data/pilot channels.

Potential PLL Discriminators

A QPSK signal \( S \) after the front-end can be modeled as:

\[
S(t) = \frac{A}{\sqrt{2}} C_D(t - \tau)D(t - \tau) \cos(2f_{IF}t + \phi) + \frac{A}{\sqrt{2}} C_p(t - \tau) \sin(2f_{IF}t + \phi) + n(t)
\]

where

- \( A \) is the amplitude associated with the total (data and pilot) signal power \( P \) (\( A = \sqrt{2P} \)),
- \( C_D \) and \( C_p \) are the spreading codes for the data and pilot channels respectively,
- \( f_{IF} \) is the Intermediate Frequency (IF)
- \( \tau \) and \( \phi \) are the incoming signal code delay and carrier phase, and
- \( n \) is the filtered Gaussian noise.

Assuming a local phase estimate \( \hat{\phi} \), and a code delay estimate \( \hat{\tau} \), the carrier wipe-off and the correlation with the locally generated code on each channel will lead to:
\[ I_D(t) = \frac{A}{2\sqrt{2}} R_D(\varepsilon, \tau) D(t - \tau) \sin c(\pi f T) \cos(2 \varepsilon) + n_{i_D}(t) \]

\[ Q_D(t) = \frac{A}{2\sqrt{2}} R_D(\varepsilon, \tau) D(t - \tau) \sin c(\pi f T) \sin(2 \varepsilon) + n_{q_D}(t) \]

\[ I_p(t) = \frac{A}{2\sqrt{2}} R_p(\varepsilon, \tau) \sin c(\pi f T) \cos(2 \varepsilon) + n_{i_p}(t) \]

\[ Q_p(t) = \frac{A}{2\sqrt{2}} R_p(\varepsilon, \tau) \sin c(\pi f T) \sin(2 \varepsilon) + n_{q_p}(t) \]

where

- \( \varepsilon \) and \( \varepsilon \) are the errors in phase and code delay estimation respectively,
- \( \Delta f \) is the error in Doppler estimation,
- \( T \) is the pre-detection integration time,
- \( R_D \) and \( R_p \) are the normalized correlation functions associated with the codes \( C_I \) and \( C_Q \) respectively,
- \( n_{i_D}(t) \), \( n_{q_D}(t) \), \( n_{i_p}(t) \), \( n_{q_p}(t) \) are uncorrelated Gaussian noises with a variance of \( \frac{N_0}{4T} \).
- \( N_0 \) is the power spectrum density of the thermal noise.

Due to the presence of navigation bits on the data channel, the discriminator associated with this channel has to be invariant to bit transitions. As a consequence, a Costas loop has to be used, resulting in non-optimal tracking with degradation in terms of noise due to squaring and to the operational range. The traditional Costas discriminator is given by (neglecting the Doppler error):

\[ D_{\text{Data}} = I_D Q_D = \frac{A^2}{16} R_D^2(\varepsilon) \sin c(2 \varepsilon) \] (3)

The drawbacks of this discriminator are twofold: (1) it uses squaring, (2) it requires normalization, achieved using \( I_D^2 + Q_D^2 \), that will bring extra noise. Its operational range is \( [-\pi/2, \pi/2] \) in theory. However, this Costas discriminator actually tracks twice the phase error, and in practice has an operational range of only \( [-\pi/4, \pi/4] \).

Another widely used discriminator for the data channel is the classical arctangent discriminator. It is obtained by:

\[ D_{\text{Data}} = \arctan \left( \frac{Q_D}{I_D} \right) = \varepsilon \] (4)

This discriminator has the advantage of not needing to be normalized. However, the presence of uncorrelated noise on the numerator and denominator degrades its estimation, especially at low C/N0. Its operational range is wider than the Costas discriminator, and equals \( \left[-\frac{\pi}{2}, \frac{\pi}{2}\right] \).

When looking at the pilot channel, because no data is present, it is possible to use adapted coherent discriminators. One interesting discriminator, that will be referred to as the coherent discriminator, was proposed by Hegarty (1999) and has the following expression:

\[ D_{\text{Pilot}} = Q_p = \frac{A}{2\sqrt{2}} R_p(\varepsilon) \sin c(\varepsilon) \] (5)

This discriminator has the advantage not having squaring losses, which could be very helpful at low C/N0. It also requires normalization, using \( \sqrt{I_D^2 + Q_D^2} \) for instance.

Its operational range is \( \left[-\pi, \pi\right] \).

Finally, a second discriminator mentioned in the literature (Macabiau et al., 2003) is the extended (or four-quadrant) arctangent discriminator:

\[ D_{\text{Pilot}} = \arctan 2(Q_D, I_D) = \varepsilon \] (6)

Its operational range is \( \left[-\pi, \pi\right] \), twice as large as the traditional arctangent discriminator. In terms of resistance to thermal noise, it has the same performance as the classical arctangent discriminator for high C/N0, as it equals it for errors within \( \left[-\frac{\pi}{2}, \frac{\pi}{2}\right] \). Moreover, unlike the other discriminators presented, it can track the phase modulo \( 2\pi \), so without a half a cycle ambiguity.

These four discriminators are all candidates to be used on the data/pilot receiver described herein. They were tested for comparison for three C/N0, with a usual PLL loop bandwidth of 10 Hz, and an integration time of 1 ms. The results are given in Table 1. For high C/N0, the difference between the different discriminators is minor. However, for lower C/N0, the coherent discriminator seems to perform the best due to its pure integration (no squaring). However, one has to be careful with these standard deviations. Indeed, when the operational range of a discriminator is narrow, the output of the discriminator, whatever the noise level, will remain within this range, but with a cycle slip. As a consequence, the resulting discriminator output might not correspond to the actual stress, and it corrupts the statistics when only noise is considered. A second observation concerns the use of \( I_D^2 + Q_D^2 \) as normalization. This normalization follows a non-central \( \chi^2 \)-square distribution, and has a mean value equal to \( \frac{A^2}{8} + 2 \text{Var}(n_i) \) that will overestimate the signal.
power and will underestimate the phase error as a consequence. This is particularly true for low C/N0 and could impair carrier tracking.

It is important to also mention that the width of the discriminator’s operational range has a significant impact on its resistance to dynamic stress. A wider range will enable a correct reaction to higher dynamics before undergoing a cycle slip. The extended arctangent discriminator has the widest operational range, and the classical arctangent discriminator has one equivalent to the coherent discriminator. Moreover, they do not require normalization. This makes them very suitable for easy implementation for carrier tracking using the data/pilot signals.

Table 1 – Standard Deviation of the Four PLL Discriminators for a C/No of 30, 35 and 40 dB-Hz, a Pre-Detection Time of 1 ms and a Loop Bandwidth of 10 Hz In Presence of Thermal Noise Only (radians)

<table>
<thead>
<tr>
<th>Discriminator</th>
<th>30 dB-Hz</th>
<th>35 dB-Hz</th>
<th>40 dB-Hz</th>
</tr>
</thead>
<tbody>
<tr>
<td>Norm. $I_D Q_D$</td>
<td>0.122</td>
<td>0.065</td>
<td>0.037</td>
</tr>
<tr>
<td>Norm. $Q_P$</td>
<td>0.1045</td>
<td>0.065</td>
<td>0.039</td>
</tr>
<tr>
<td>$\arctan\left(\frac{Q_D}{I_D}\right)$</td>
<td>0.147</td>
<td>0.071</td>
<td>0.039</td>
</tr>
<tr>
<td>$\arctan\left(2\frac{Q_D}{I_D}\right)$</td>
<td>0.134</td>
<td>0.073</td>
<td>0.039</td>
</tr>
</tbody>
</table>

Now that the choice of the discriminator for the data and pilot channels has been investigated, the design of the PLL loop filter can be discussed to accommodate the different channel characteristics. The different sources of error will have different impacts in the design of the loop filter. They are studied in the following sub-section.

PLL Error Sources

An extensive study of carrier tracking sources of error can be found in Kaplan [1996], Van Dierendonck [1997], Hegarty [1997], and Irsigler and Eissfeller [2001]. The following discussion summarizes the main results.

It is well-known that the impact of thermal noise on carrier phase tracking using a Costas discriminator is given by [Kaplan, 1996]:

$$\sigma^2_{PLL,Noise} = \frac{B_L}{C} \left(1 + \frac{1}{2T} \frac{C}{N_0}\right) \text{(rad}^2\text{)}$$  \hspace{1cm} (7)

where $B_L$ is the loop bandwidth in Hz.

For the coherent discriminator, because there is no squaring occurring, the term between brackets can be eliminated, giving better performance.

The receiver oscillator creates a phase jitter affecting the PLL VCO due to slight instabilities of its central frequency. This can seriously impair carrier tracking due to potential sudden variations that are dependent upon the quality of the oscillator. The oscillator frequency noise power spectral density (PSD) can be expressed by:

$$S_{Osc,Err}(f) = \frac{h_{-2}}{2f^2} + \frac{h_{-1}}{2f} + \frac{h_0}{2}$$  \hspace{1cm} (8)

where $h_{-2}$, $h_{-1}$, $h_0$ represent the random walk, Flicker and white components of the oscillator frequency error. Typical values are given in Winkel [2003].

The phase jitter induced by the oscillator phase error has the following properties: (1) it decreases when the loop bandwidth increases; and (2) it increases with the order of the loop filter. The following expression [Irsigler and Eissfeller; 2001] represent the variance due to the oscillator phase error for a third order loop filter.

$$\sigma^2_{PLL, Osc} = 2\pi^2 f_{Osc}^2 \left(\frac{\pi^2 h_{-2}}{3\omega_L^2} + \frac{\pi h_{-1}}{3\sqrt{3}\omega_L^2} + \frac{h_0}{6\omega_L^2}\right) \text{(rad}^2\text{)}$$  \hspace{1cm} (9)

where $\omega_L = 1.2B_L$ (for a third order PLL).

The oscillator phase can also be disturbed by vibrations. The PSD of the vibrations varies according to the user and the application. Several examples can be found in Hegarty [1997] and Irsigler and Eissfeller [2001]. The phase estimation error due to oscillator vibration possesses the same characteristics as that due to the oscillator frequency noise: it increases with the loop order and decreases with the loop bandwidth. For a flat vibration PSD between the pulsations $\omega_1$ and $\omega_2$ and null elsewhere, and considering a third order loop, the standard deviation is given by [Irsigler and Eissfeller; 2001]:

$$\sigma^2_{PLL, vib} = \frac{2\pi^2 f_{Osc}^2 k_g^2 G_h}{\omega_L} \times$$  \hspace{1cm} (10)

$$\left[\frac{1}{3} \left(\arctan\left(\frac{\omega_2}{\omega_1}\right) - \arctan\left(\frac{\omega_1}{\omega_2}\right)\right) + \frac{1}{6} \arctan\left(\frac{\omega_1 \sqrt{3} + 2\omega_2}{\omega_2}\right) - \arctan\left(\frac{\omega_2 \sqrt{3} + 2\omega_1}{\omega_1}\right)\right]$$

$$\left(\text{rad}^2\right)$$

where $k_g$ is the oscillator’s g-sensitivity in parts-per-g, and

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\( G_g \) is the single-sided vibration power spectral density assumed constant between \( \omega_1 \) and \( \omega_2 \), and null elsewhere.

Finally, the usual phase bias due to dynamics can be found in Kaplan [1996]. For a third order loop filter, it is equal to:

\[
\theta = \frac{\pi}{180} \frac{dR^3}{\omega_i^3} \text{ (rad)}
\]

(11)

where \( R \) is the line-of-sight distance between the satellite and the receiver antenna (in metres).

As seen, the mitigation of the thermal noise requires a narrow loop bandwidth, while the other sources of error require a large loop bandwidth. This is why these four sources of error should all be taken into account when designing a PLL loop filter. They are represented in Figure 1 for a C/No of 30 dB-Hz, a TCXO oscillator (characteristics in Winkel [2003]), vibration taken from Irsigler and Eissfeller [2001], an integration time of 20 ms, and for a third order Costas PLL with a loop bandwidth ranging from 2 to 20 Hz. It shows that the choice of the loop bandwidth should be realized according to previous knowledge of the expected errors. The relative magnitude of each error has to be assessed to design a 'worst case' scenario that can be supported by the PLL.

\[
\frac{\pi}{2B_L \gamma} \tanh \left( \frac{2 \pi \gamma}{\sigma_{PLL}} \right) \times \sum_{n=1}^{\infty} \left( -1 \right)^n \frac{I_n \left( \frac{1}{\sigma_{PLL}} \right)}{1 + \left( \frac{n \sigma_{PLL}}{\gamma} \right)^2}
\]

where \( \gamma \) is the constant phase tracking error due to dynamics (in radians), and \( I_n \) is the n-th order modified Bessel function of the first kind.

For a third order loop, an additional 2 dB in the C/No is required to achieve the same performance as a first order loop [Hegarty, 1997]. It also has to be noted that when using a Costas discriminator, the value of \( \sigma_{PLL} \) and \( \gamma \) have to be doubled in order to take into account that it actually tracks twice the phase [Hegarty, 1997]. Figure 2 shows the probability of a cycle slip within one second as a function of the loop bandwidth for different scenarios for the Costas discriminator, and the coherent discriminator. Although the figure is theoretically not correct (as the sources of error are not white noise), the methodology used here follows Hegarty [1997] and aims to show the level of cycle slips to be expected. It is obvious that the use of the Costas discriminator would seriously impair carrier tracking compared to other discriminators that have a wider discriminator range.

\[
P_{\text{cs}} = 1 - e^{-\frac{T_{\text{cs}}}{T}}
\]

(12)

with

\[
\frac{\pi}{2B_L \gamma} \tanh \left( \frac{2 \pi \gamma}{\sigma_{PLL}} \right) \times \sum_{n=1}^{\infty} \left( -1 \right)^n \frac{I_n \left( \frac{1}{\sigma_{PLL}} \right)}{1 + \left( \frac{n \sigma_{PLL}}{\gamma} \right)^2}
\]

Figure 1 – PLL Errors Standard Deviation for a C/No of 30 dB-Hz (Top)

In order to have a better idea on the loop bandwidth to choose, it is important to look at another figure of merit characterising the PLL performance: its resistance to cycle slips. It constitutes an intuitive mean to evaluate the performance of the PLL. Holmes (1990) evaluates the probability of having a cycle slip within a period of time \( T_{\text{cs}} \) for a first order PLL, assuming a flat noise PSD, as:

\[
0.01
\]

Figure 2 - Probability of Cycle Slip in One Second for a 25 dB-Hz C/No considering All Sources of Error and a Jerk of 1 g/s

On the other hand, the use of the coherent discriminator, or even of the classical arctangent discriminator (that has the same operational range, but more thermal noise), would give better results. The extended arctangent discriminator, because of its wider operational range,
would even resist further dynamic stress and a low $C/N_0$. From Figure 2, it is also noticeable that the choice of the loop bandwidth is very dependent upon the level of errors expected, and that it should be chosen according to the worst case scenario.

Now that the different sources of error affecting carrier tracking have been discussed, and that the differences between potential discriminators have been outlined, the design of the data/pilot carrier tracking can be presented.

**Data/Pilot Carrier Tracking Proposed Implementation**

Since the data and pilot discriminators estimate the same phase error, they can be combined to produce more consistent phase estimation. The first method using a combination of data and pilot channels was proposed by Hegarty [1999] and Tran and Hegarty [2002]. The two discriminators’ output noise can be considered as independent due to the different spreading codes used on each channel. As a result, it is possible to reduce the noise variance of the estimate through the combination of the two discriminators’ outputs. For this research, the extended and classical arctangent discriminators were selected for the pilot and data channels due to their large operational range, allowing better resistance to dynamic stress, and because they behave the same in the range $[-\pi/2; \pi/2]$, easing their combination. The combined discriminator proposed by Hegarty (1999) can be written as:

$$D_{comb} = \alpha D_{data} + \beta D_{pilot}$$

(14)

In order to have an optimal estimation process, $\alpha$ and $\beta$ should be chosen as:

$$\alpha = \frac{\sigma^2_{D_{pilot}}}{\sigma^2_{D_{data}} + \sigma^2_{D_{pilot}}}; \beta = \frac{\sigma^2_{D_{data}}}{\sigma^2_{D_{data}} + \sigma^2_{D_{pilot}}}$$

(15)

For high $C/N_0$, and assuming a common pre-detection integration, the two discriminators have the exact same phase error variances when only thermal noise is considered, as already mentioned. For low $C/N_0$, however, the smallest operational range of the classical arctangent discriminator could degrade the combined estimation process. Furthermore, as it has been seen, the behaviour of the two discriminators will also be different when high dynamics (including oscillator phase error and vibration) is taken into account. When looking at the cycle slip issue for this combined estimation method, the PLL loop filter and the phase loss of lock detector have to be designed according to the data channel discriminator as it is the most vulnerable. Indeed, a cycle slip in the data discriminator will result in a cycle slip in the combined discriminator. This might be a drawback as it is not possible to fully exploit all the possibilities offered by the wide operational range of the extended arctangent discriminator. In order to compensate for this, one might want to use only the pilot channel if sudden high dynamics are detected. This can be done when directly checking at the output of the discriminator. The new proposed method takes this fact into account. If a cycle slip occurs on the data channel, but not on the pilot channel, the difference in the estimate of the phase error will suddenly jump significantly. As a result, by constantly checking the difference between the output of the data and pilot discriminators, one could decide between using both, to have better noise mitigation, or just the pilot channel, to have better resistance to dynamics. Moreover, it still allows for a reduction in the loop bandwidth according to the extended arctangent discriminator as, even if the loop bandwidth is too narrow for the data channel discriminator, it will still rely on the pilot channel. The method proposed herein can be modeled by:

$$D_{comb} = \frac{D_{data} + D_{pilot}}{2} \quad \text{if } |D_{data} - D_{pilot}| \leq \eta$$

$$D_{comb} = D_{pilot} \quad \text{else}$$

(16)

The choice of $\eta$ is particularly important, as the smaller it is, the more the data channel will be excluded, and the larger it is, the better the thermal noise will be mitigated. A relatively tight value of $\eta = \frac{\pi}{4}$ has been chosen for the rest of this paper, as it has been estimated that it is more important to have reliable tracking, even if noisier, than to have better noise mitigation, but a higher chance of losing lock.

While this method can be near optimal for high $C/N_0$, it might have problems for low $C/N_0$ as the thermal noise component is far more important. Because of the tight choice of the $\eta$ value, the method will follow the pilot channel, and should not be affected by early loss of lock due to cycle slips. Figure 3 shows an example of the use of this new method. The $C/N_0$ is 45 dB-Hz, the loop bandwidth has been set to 5 Hz, and the integration time to 1 ms. The loop bandwidth has been chosen very narrow on purpose to test for the resistance to cycle slips. In order to simulate a sudden change in phase, a low-cost oscillator has been modeled (quartz) for the satellite to generate sudden small range changes. Because the tracking starts with a $\pi$ difference between the data and pilot channels (the classical arctangent tracks the carrier phase modulo $\pi$), the combined discriminator will follow the pilot channel that tracks the true phase, as expected. After 310 ms, a sudden change in the incoming phase creates a cycle slip on the data channel that is not present on the pilot channel. The data/pilot combined discriminator follows, as desired, the pilot channel without experiencing any cycle slips either. In addition, because the phase is better tracked, the data and pilot
channels are used most of the time in the new algorithm (> 95% of the time in this example), resulting in better noise mitigation. This can be seen through the fact that the data-pilot curve (blue) is thinner than the pilot-only (red). More results are given in the last section of this paper.

Figure 3 – Instantaneous Doppler Estimate for a C/No of 45 dB-Hz and High Dynamics

This proposed method reduces the thermal noise affecting carrier phase estimation, while it is still able to track the signal’s dynamics in a robust manner. It will allow better aiding for code tracking if carrier aiding is used, and it will allow for longer coherent integrations, as the Doppler estimate is also more accurate. This can be a great advantage for the code tracking loop as well as for techniques demanding long integration times, like most of the efficient multipath mitigation techniques.

CODE TRACKING

The sources of errors for code tracking are similar as those for the phase. However, they will not have the same impact. For instance, the oscillator phase error and vibrations will have a negligible impact on code tracking due to their limited magnitude. A significant source of error is multipath, especially for signals with low code chip rates, which are the focus of this study. Many methods have been formulated in order to mitigate multipath-induced errors. Two methods will be studied herein that will lead to different implementations when used in a data/pilot receiver: the Narrow Correlator™ technique, and the High Resolution Correlator.

Narrow Correlator (NC)
The Narrow Correlator (NC) technique [Van Dierendonck et al., 1992] is well known. It uses the conventional early-late tracking method, but using a narrower spacing between the early and late correlators, allowing for better noise mitigation. It also brings stronger multipath mitigation compared to wide correlator spacing. However, while it succeeds in mitigating multipath, it does not eliminate it, and strong multipath with a delay anywhere between 0 and slightly more than 1 chip can impact significantly the estimation of the code delay estimate. The discriminator that will be used in this section is the following:

\[ D_{NC} = \frac{I_{E-L}I_p + Q_{E-L}Q_p}{I_p^2 + Q_p^2} \]  

(17)

High Resolution Correlator (HRC)
This method is described in details in McGraw and Braasch [1999]. It approximates tracking using a locally generated second derivative of the spreading code described by Weill (1997). This locally generated code is shown in Figure 4 and can be modeled as:

\[ C_{HRC}(t) = 2 \times c(t) - c(t + d) - c(t - d) \]

where

\[ c(t) \]

is the spreading code used in the incoming signal, while \( d \) is the parameter defining the HRC method.

Figure 4 – Example of Local Code Replica for BOC(1,1) Tracking Using NC and HRC

As seen in Figure 4, this method consists in blanking a great part of the incoming code. This will have a great impact on the mitigation of multipath, as they will be blanked if there delay is greater than \( 2d \). However, this blanking will also create a strong degradation of the correlator output. As an example, for a BPSK(1) signal using the HRC method with \( d = 0.05 \) chips, the equivalent degradation in C/No is 13 dB, as shown in Figure 5. This figure shows the correlation function between the incoming signal and the local code using the HRC technique, and the NC method for a BOC(1,1) and a BPSK(1) signal. The discriminator used by McGraw and Braasch (1997) is:

\[ D_{HRC}^1 = \frac{I_{E-L}^2 I_p + Q_{E-L}^2 Q_p}{I_p^2 + Q_p^2} \]

(18)

The spacing between the early and late correlators is equal to \( 2d \).
This discriminator has the advantage to use the correlator prompt values obtained by the traditional correlation of the incoming signal with the actual spreading code, resulting in tracking which is only slightly degraded (code delay error variance increased by 3 dB compared with a NC technique for a BPSK(1) using the same early-late spacing [McGraw and Braasch, 1999]). Its resistance to multipath is shown in Figure 6 along with the resistance of the NC technique for the BPSK(1) and BOC(1,1) signals using a front end filter of 24 MHz. It has to be noted that because narrow correlator spacing is desired to efficiently remove multipath, large front end filter bandwidths are also required. The improvement brought by the HRC technique is significant for medium multipath delays. It appears to be less optimal for BOC(1,1) signals due to the two lobes around 0.5 chips, although improving drastically over the NC technique.

One main drawback of the HRC discriminator is its limited tracking domain, as shown in Figure 7. The shape of its discriminator output is not well suited for tracking as it fades away when the code delay error reaches \( d \) chips. This is one of the reasons why it might not be suited for low C/No or for tracking without carrier aiding. In the method presented herein, it is still beneficial with respect to other derivations of the HRC discriminator (using the degraded ‘HRC’ prompt correlator) that would undergo a much higher noise. The reason for this will be explained in the next sub-section.

Figure 5 – Correlation Function used by the HRC Technique (\( d = 0.05 \) chips) and by the NC Technique for a BOC(1,1) (Top) and a BPSK(1) (Bottom) Signal

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Figure 6 – NC and HRC (\( d = 0.05 \) chips) Multipath Envelopes for BPSK(1) and BOC(1,1) Signals Using a 24 MHz (Double-Sided) Front-End Filter

Data/Pilot Code Tracking Implementation

When considering the use of two channels tracking signals having the same phase and code delay, one can think of many possible implementations. As seen in the study of carrier phase tracking, the use of the extended arctangent is tremendously promising, but, to be optimal, it would require prompt values as clean as possible. This means that the prompt values given by the pilot channels will have to use prompt values of the same type as the one given by the NC approach. As shown previously, this is the case for the HRC technique.

The narrow correlator technique is more robust in terms of the stability domain and thermal noise mitigation. Moreover, in order to drive the carrier tracking as well as possible, reliable code tracking is needed. This makes the NC technology the main candidate for the pilot channel tracking technique. However, reserving the data channel for the HRC method would cancel the opportunity to improve drastically its correlation gain through longer coherent integrations to get a better discriminator output. One has to remember that the use of an extended arctangent makes the carrier tracking far more robust, as seen in the previous section, and that carrier-aiding will be provided more reliably, ensuring the use of a low DLL loop bandwidth, and so a reduction in the code delay error in the presence of thermal noise. The risk of having a strong stress that could not be compensated by the HRC discriminator, although real, is then significantly reduced.
It is then possible to have a DLL on the pilot channel using the HRC technique which is reliable.

However, the risk of losing code lock still remains due to the limited HRC discriminator linear domain. From McGraw and Braasch [1999], Figure 8 shows the standard deviation of the discriminator output in the presence of thermal noise for a BPSK(1) signal for different coherent integration times using the HRC technique with \(d = 0.05\) chips. If one wants to limit the noise error on the discriminator output to 0.05 chips (\(3\sigma\)), to make sure that the code delay error due to thermal noise be within the HRC tracking region, and if the receiver is expected to work safely until 25 dB-Hz, a coherent integration time longer than 500 ms would be required, which might be too long for certain applications.

\[\text{Figure 8 – Discriminator Output Standard Deviation for the HRC Technique for a BPSK(1) Signal, Coherent Integrations of 20, 50, 100, 200, 300, and 500 ms and } d = 0.05 \text{ chips}\]

To reduce the chance of losing lock early on the pilot channel, one also has to think that there is access to the data channel as well. The data channel can be used as a ‘guardian’ to ensure that the code error remains within the tracking domain of the HRC discriminator. If the data pilot detects a difference in the code delay estimation error reaching \(d\) chips, the DLL tracking loop could switch to the more robust NC tracking. The HRC discriminator output would be controlled through the data channel discriminator (using an NC technique). Based on the same principle as for the carrier tracking channel, the discriminator used will be dependent upon the difference between the data and pilot channel discriminator output:

\[D_{\text{comb}} = D_{\text{data}} \quad \text{if} \quad \left( \left( \text{sgn}(D_{\text{data}}) = \text{sgn}(D_{\text{pilot}}) \right) \& \left| D_{\text{data}} - D_{\text{pilot}} \right| > \xi \right) \]

\[\text{or} \quad \left( \left( \text{sgn}(D_{\text{data}}) = -\text{sgn}(D_{\text{pilot}}) \right) \& \left| D_{\text{data}} - D_{\text{pilot}} \right| > \kappa \right) \]

\[D_{\text{comb}} = D_{\text{pilot}} \quad \text{else} \quad (19)\]

The choice of \(\xi\) and \(\kappa\) can be dependent upon the C/No value, or according to the front end filter bandwidth (that will transpose in the shape of the discriminator shown in Figure 7). Such an implementation will bring robustness to the implementation, as it should be as robust as the NC technique. When higher C/No is available, strong and robust multipath mitigation can then be achieved. For lower C/No, although the multipath mitigation might not be as efficient, tracking will remain robust. Note that no real combination of the data and pilot discriminators is done. The reason is that combining them would cancel some of the benefits brought by the HRC technique in mitigating multipath, as the multipath error would enter the code delay estimate through the NC discriminator. It should be noted that it is not necessary to have the same integration times (coherent and/or non-coherent) on both channels as long as they have a common divider. However, if a different integration time for each channel is chosen, the algorithm will have to be slightly modified.

**SIMULATION RESULTS**

Now that two methods using data and pilot channels for carrier phase and code delay tracking have been proposed, simulations were undertaken to verify their behavior and to assess their performance. Several tests were realized to compare the performance of the new method with respect to other traditional methods. In order to respect the traditional digital loop filter theory, the coherent integration time on the pilot channel is limited to \(T < \frac{1}{100B_L}\). Although a value smaller than 100 could be taken, this constraint was set to make sure the loop filters were behaving correctly. Simulation of the incoming signal was made through the use of a GNSS software IF signal generator developed by the PLAN group at the University of Calgary. It allows flexibility in the conception of fairly complex scenarios [Julien et al., 2004]. The tests simulated one Galileo satellite transmitting a QPSK BOC(1,1) signal. The data rate on the data channel was set to 250 sps, allowing coherent integration of only 4 ms. No tiered code were simulated. The codes used were taken from the C/A codes used by GPS. The IF frequency was set to 7 MHz, and the double-sided front-end filter bandwidth was set to 12 MHz. Although this bandwidth is not optimal for use of the HRC technique, it still allows an improvement that should be visible on the results (and it saves a lot of computation time as it allows a lower sampling frequency).

Two different tests were undertaken in order to test the new data/pilot technique performance, and to compare it against other classical single channel techniques. The first one was designed to test for the reaction of the code tracking loop to a large stress, the second one was...
designed to test the actual performance of the proposed technique in the presence of multipath.

Loops Reaction to a Large Stress
This scenario was designed to test the reaction of the new tracking method to a large stress. As a result, tracking was started with a code delay error of 0.126 chips in order to see the reaction of the DLL. Moreover, a Quartz clock was simulated in the satellite in order to add some extra-dynamics to the signal. This allows testing for the response of the PLL to dynamic stress as well. The total signal C/N$_0$ was 40 dB-Hz (it then goes through filtering and 1-bit quantization). The receiver parameters were set in the following way:
- DLL: Loop bandwidth of 1/3 Hz, HRC technology using a coherent integration of 28 ms for the pilot channel, and NC technology using 7 non-coherent integrations of 4 ms on the data channel. The early-late spacing was set to 0.1 chips. Carrier aiding is used
- PLL: Loop bandwidth of 8Hz, coherent integration time of 4 ms on both channels. The loop bandwidth was chosen tight on purpose in order to test the capacity to resist to cycle slip.

For the proposed data/pilot code tracking method, the parameters $\xi$ and $\kappa$ were set to 0.04 and 0.08 chips respectively. The receiver antenna was assumed static. Figure 9 shows the response of the DLL to the initial code delay error for the data-only channel using the NC technique, the pilot-only channel using the HRC technique, and the proposed data/pilot technique. The data-only configuration has the best response, as it uses the NC technology which is the more robust among the tracking methods used herein. The pilot-only channel, because it uses the HRC technique, has a slower response to the step error, as expected, due to its discriminator shape. The proposed data/pilot configuration has a behavior in between both, but far closer to the NC response. This means that for large errors, it is able to quickly recover compared to using the HRC technique alone.

Figure 10 shows the estimation of the Doppler from the PLL. Many cycle slips occur on the data-only channel using a classical arctangent discriminator, as shown by the very ‘noisy’ Doppler estimate. There is almost no correct carrier tracking at all. Although a few cycle slips are present for the two other configurations, they are far less numerous and show a better behavior of the PLL driven by the extended arctangent (the Doppler changes rapidly due to the simulated Quartz clock). The data/pilot channel shows approximately the same performance as the pilot channel, as expected, because its PLL discriminator will follow the pilot channel discriminator in the case of high dynamics. The data/pilot implementation seems also to remove part of the noise present in the pilot-only configuration.

Test of Multipath Mitigation Capability
The second series of tests was designed to assess the ability of the new technique to mitigate multipath. Two scenarios were designed for this purpose. The first test simulates a static receiver that receives a strong specular multipath with very little phase change. This should magnify the impact of the multipath on tracking, and allows for a matching view of the multipath mitigation capacity of the new technique. Several multipath signals were simulated. One is from a large flat obstacle situated 50 meters from the antenna, and reflects the incoming signal with an amplitude of half the direct signal. Two multipath signals are ground reflections with amplitudes being one third of the incoming signal (coming from the satellite and from the obstacle). Ten ‘scattered’ multipath signals that have a power at least 16 dB lower than the direct signal are simulated on the same basis as Hegarty [2004]. Two C/N$_0$ (total data+pilot) were simulated: 50 and 40 dB-Hz at the antenna level. The signal then goes through a 12 MHz double sided bandwidth filter and a 1-
bit quantization, further reducing the signal-to-noise ratio. A receiver oscillator phase jitter is simulated using the model given by Winkel [2003]. The oscillator simulated is a TCXO. The second test has exactly the same configuration, except that the receiver is describing a 50 radius circle just beside the same obstacle at a speed of 100 km/h.

When the simulated C/N_0 at the antenna level was set to 50 dB-Hz, the DLL loop bandwidth was set to 1 Hz with a coherent integration time of 8 ms for the pilot channel and two non-coherent integrations of 4 ms for the data channel. This was intended to make the receiver more susceptible to multipath error. For the 40 dB-Hz test, the exact same configuration as for the scenario testing the resistance to a stress was used.

The three methods compared were the described data/pilot technique, the HRC technique on the pilot channel-only, and the NC technique on the pilot channel only, in order to have the same correlation gain. For the proposed data/pilot method, the parameters  \( \xi \) and  \( \kappa \) were again set to 0.04 and 0.08 chips respectively.

Figure 11 shows the code delay residual estimation result in the static case for the high and low C/N_0 scenarios. The NC technique shows a typical behaviour when the multipath delay is slowly changing. The code delay has maximum and minimum errors reaching slightly less than 0.04 chips which is typical for the type of filter used (12 MHz double-sided) and the multipath delay (around 48 metres). As expected, the HRC technique mitigates most of the error due to the presence of multipath. A small variation is visible, in line with the expected error. The proposed data/pilot method shows behavior very similar to the HRC-alone method. However, after approximately 8 seconds, a slight divergence is noticed. This is due to the fact that the error between the NC discriminator and the HRC discriminator reaches 0.04 chips, which is the value of the parameter  \( \alpha \). This can be easily calibrated, redefining a higher  \( \alpha \), to be removed. However, it is important to know that this kind of error can appear if the discriminator filter is not thoroughly designed. The resulting estimation error is very small (< 0.005 chips) compared to the multipath-induced NC error.

Figure 12 shows the results of the Doppler residual estimate. The effect of the receiver clock error is visible, as the Doppler estimate is not continuously decreasing as it should in the ideal case (the simulated satellite is raising). The proposed data/pilot technique shows a thinner line, implying better noise mitigation due to the combination of the carrier-phase discriminators. This is visible for both the high and low C/No cases. This is very important as it shows that even at fairly low C/N_0, this method helps to mitigate part of the thermal noise, as well as tracking a signal with high dynamics.

Figure 13 shows the code delay residual estimation result in the dynamic case. Because the simulated user was moving fairly fast (100 km/h) on a small circle, the multipath environment is changing quickly. As a result, multipath will have less impact on the final measurement error. This is the case when looking at the figure. Concerning the high C/No case, two spikes correspond to the moment when the user is at the furthest location from the obstacle, so when the multipath delay is changing at a slower rate, they are visible when using the NC technique. This is because the DLL loop bandwidth in the high C/N_0 case was chosen large (1 Hz), and as a result, it can better track the error resulting from the presence of multipath. These two spikes are not visible on the proposed data/pilot and HRC methods. For the low C/N_0 case, because the DLL loop bandwidth has been chosen to be small (1/3 Hz), the two spikes are not visible due to the slower response of the DLL loop filter. In such a case, the NC technique shows a slightly better behavior essentially due to a better mitigation of the thermal noise (around 3 dB).
CONCLUSIONS

A new method to mitigate carrier phase noise and code multipath has been proposed using both the data and pilot channels that will be available in most of the future GNSS signals. This method has two components. It concerns carrier tracking first, providing robust tracking driven by an extended arctangent discriminator on the pilot channel. It is aided, in the case of low to medium dynamics, by a classical arctangent discriminator on the data channel through a linear combination in order to mitigate the ambient noise. This allows having both reliable carrier phase tracking, and more accurate Doppler estimation compared to the use of the pilot channel only. The proposed method also concerns code delay tracking. The multipath resistant HRC discriminator drives the pilot channel DLL. Although it gives a noisier discriminator output than the NC discriminator, this can be compensated by the use of longer integration times that are possible due to better Doppler estimation coming from the proposed data/pilot PLL. The lack of robustness of the HRC due to its discriminator shape is compensated by the use of a more reliable carrier aiding, and the use of the NC discriminator on the data channel as a guardian of its behavior.

This method has been shown to be very effective for low and high C/N0, and for high and low dynamics in presence of strong multipath.

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